Practical Feedback Loop Analysis for Current-Mode Boost Converter

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ABSTRACT

Current-mode control is the industry standard method of controlling switching power supplies. Right-half-plane (RHP) zero expression is exactly the same as that for voltage-mode control (SLVA633). Since the LC-filter resonance is eliminated with the current feedback, there is much less phase delay in the power stage transfer function, and compensation is much easier. A Type II compensator is needed to design the loop for current-mode boost converter, and the use of the Type II compensator greatly simplifies the design process. This application report describes how to select the placement of compensation poles and zero, explaining the subharmonic oscillation phenomenon and ramp addition for slope compensation in the current-mode controlled boost converter.

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1 Introduction

Voltage-mode control, also called duty-cycle control, contains a single loop and adjusts the duty cycle directly in response to output voltage changes. Current-mode control, also called current-programmed mode or current-injected control, is a multiple-loop control method that contains two loops (an inner current loop and an outer voltage loop). There are several types of current-mode control methods, and the most popular method is fixed-frequency peak-current-mode control with fixed-slope compensation ramp. The technique is called current-mode control because the inductor current is directly controlled, whereas the output voltage is controlled only indirectly by the current loop. A control reference is used to regulate the peak current of the converter directly, simplifying the dynamics of the converter.

Figure 1 shows the schematic of the boost converter with current-mode control. As with the buck converter, the current is usually sensed in the power switch.
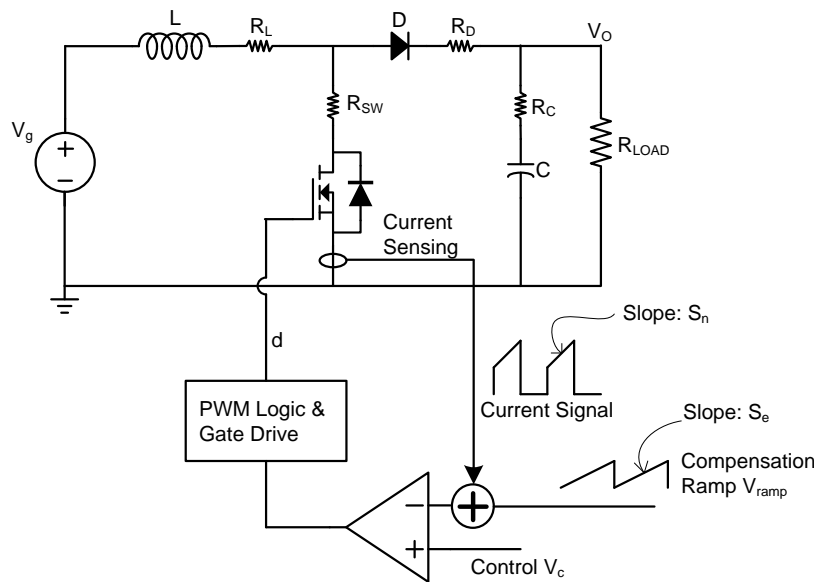

Figure 1.   Boost Converter with Current-Mode Control

Rather than using a sawtooth ramp to control the duty cycle of the converter, the simplest form of current-mode control regulates the peak of the inductor current (or switch current, depending on where the sensing is done) with a control signal, V_c. In some cases the compensation sawtooth ramp is retained to stabilize the current loop feedback, and increase noise immunity.

We typically do not sense the inductor current directly, because it is inconvenient or inefficient. The power switch current is usually sensed to gather the information about the inductor current.

2 Subharmonic Oscillation

When current-mode control was first introduced to the power electronics community in the early 1980s, it was immediately seized upon as a superior control scheme. This simple control scheme, however, had an inherent oscillation phenomenon. This is, of course, well known and documented. If you have been in power supplies for some time, you know that retaining the sawtooth compensating ramp in the control system eliminates the problem.

Figure 2 shows the nature of the current loop oscillation. This figure shows the control waveform regulating the peak current at greater than a 50% duty cycle. The steady-state waveform can exist with the clock initiating the on-time of the switch, and the control voltage terminating the on-time.
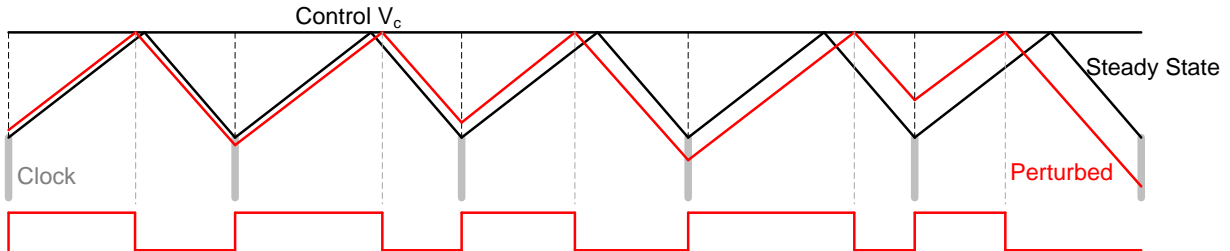

Figure 2. Subharmonic Oscillation Waveforms

In the red waveform, the inductor current is perturbed at the beginning of the cycle. This perturbation will reach the same peak current, but at the next clock cycle, the perturbation has become negative, and the amplitude has increased. After another switch cycle, the perturbation is positive again, but has increased even further.

Figure 3 shows the frequency response of current-mode boost converter without compensation ramp. Subharmonic oscillations appear as the duty cycle exceeds 50% with the following design parameters ($V_{in} = 5$ V, $V_{out} = 18$ V, $I_{out} = 3$ A, $L = 20$ µH, $F_{sw} = 200$ kHz).
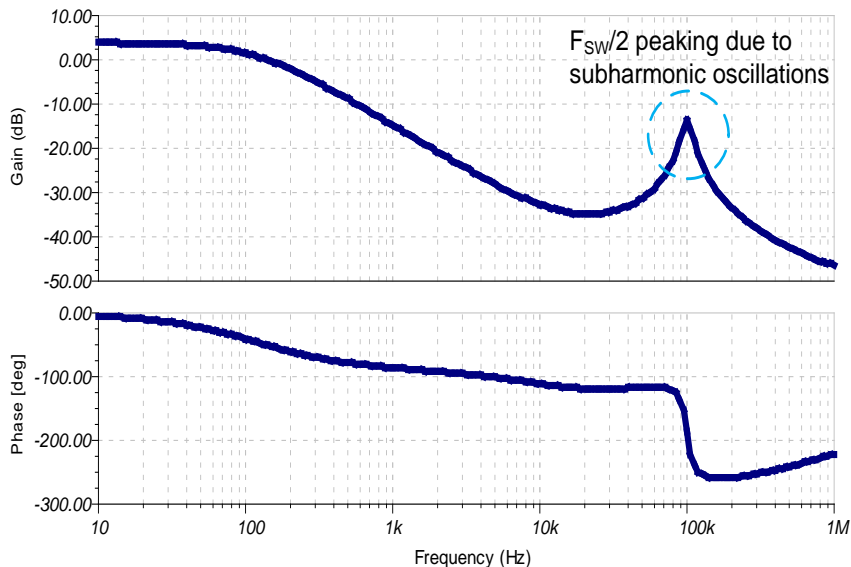

Figure 3. AC Small Signal Response without Compensation Ramp

The stabilizing effect of the compensation ramp is explained using the current feedback signal illustrated in Figure 4. The PWM waveforms are analyzed, which shows the propagation of the perturbed inductor current ($\Delta i'_L$). In the enlarged illustration in Figure 4, S_n is the slope of the on-time inductor current and S_f is the current slope of the off-time inductor current, while S_e is the slope of the compensation ramp. The ΔdT_s denotes the deviation in the on-time period due to the inductor current perturbation.
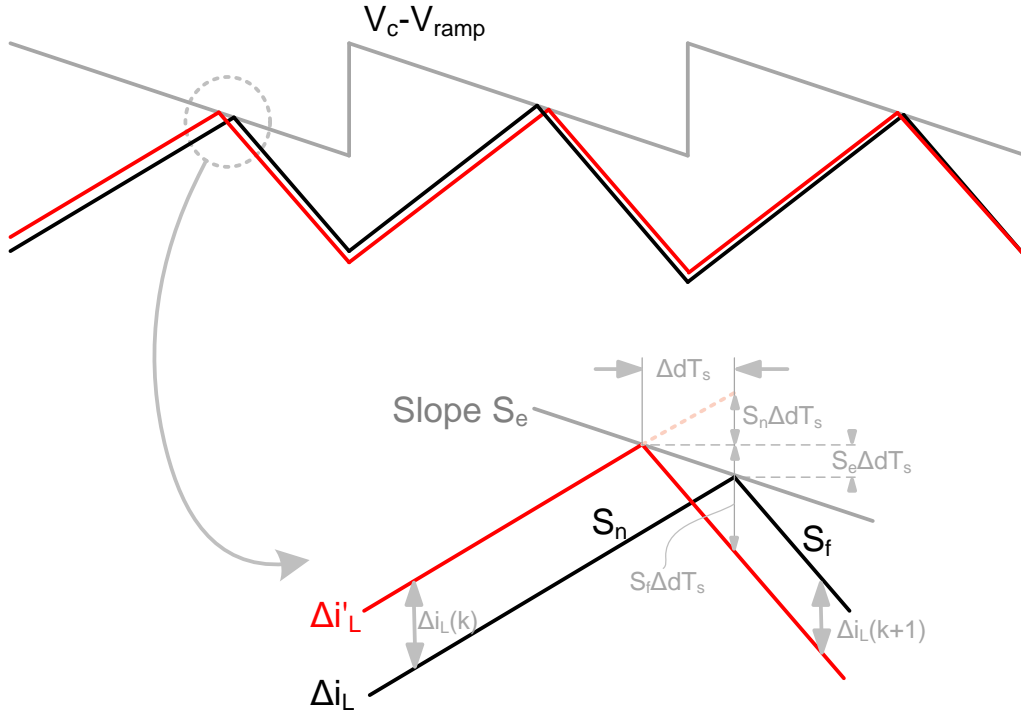
Figure 4. PWM Waveforms with Compensation Ramp

From the graphical construction, the initial distance between the original inductor current (i_L) and the perturbed inductor current (i'_L) is given by:

$$|i'_L(k) - i_L(k)| = \Delta i_L(k) = S_n \Delta dT_s + S_e \Delta dT_s \qquad (1)$$

The distance between the two currents after one operational period is given by:

$$|i'_L(k+1) - i_L(k+1)| = \Delta i_L(k+1) = S_f \Delta dT_s - S_e \Delta dT_s \qquad (2)$$

For the successive decrease in the distance between i_L and i'_L in the ensuing operational periods, the condition:

$$\frac{\Delta i_L(k+1)}{\Delta i_L(k)} = \frac{S_f - S_e}{S_n + S_e} < 1 \qquad (3)$$

is required, leading to the following condition for the compensation ramp slope:

$$S_e > \frac{S_f - S_n}{2} \qquad (4)$$

for the stabilizing effect. The exact value of the compensation ramp slope should be determined in consideration of the closed-loop performance of the converter.

Figure 5 shows the frequency response of current-mode boost converter with compensation ramp. As it is shown in Figure 5, the peaking is properly damped.
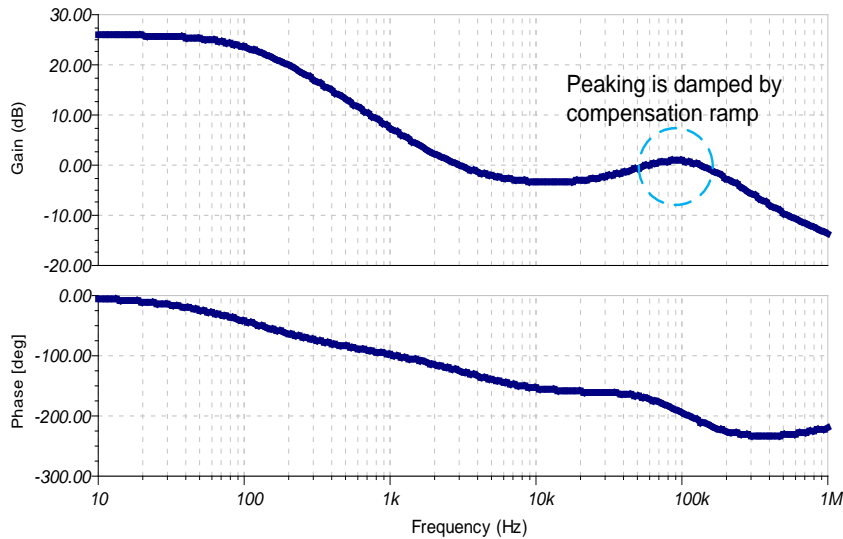
Figure 5.   AC Small Signal Response with Compensation Ramp

3   Boost Converter (Current-Mode) Transfer Function Plots

The boost converter has an additional term in the control-to-output transfer function, caused by the RHP zero of the converter:

$$\frac{\widehat{v_o}}{\widehat{v_c}} = K_{dc} \times \frac{\left(1+\frac{s}{\omega_z}\right)\times\left(1-\frac{s}{\omega_{r\Box p}}\right)}{1+\frac{s}{\omega_p}} \times f_h(s) \tag{5}$$

The dc gain of the converter is given by:

$$K_{dc} = \frac{D'\times R_{LOAD}}{R_i} \tag{6}$$

For the low-frequency part, the dominant pole is located at:

$$\omega_p = \frac{2}{C\times R_{LOAD}} \tag{7}$$

The capacitor ESR zero is at the same location as the boost converter in voltage-mode, given by:

$$\omega_z = \frac{1}{C\times R_c} \tag{8}$$

and the RHP zero is at

$$\omega_{rhp} = \frac{R_{LOAD}}{L_e} \tag{9}$$

To account for the observed oscillation in the current-mode system, the high-frequency correction term ($f_h(s)$) added to the basic power stage:

$$f_h(s) = \frac{1}{1+\frac{s}{\omega_n Q_p}+\frac{s^2}{\omega_n^2}} \tag{10}$$

Figure 6 shows the schematic of the small-signal analysis using a simple voltage-controlled voltage source as an error amplifier. On this small-signal boost, the voltage-controlled voltage source amplifies by about 89.5 dB, the difference between a portion of V_{out} and the 2.5-V reference. In order to avoid running the circuit in a closed-loop configuration, we can install an LC filter featuring an extremely low cutoff frequency.

The error amplifier can be a simple voltage-to-voltage amplification device, that is, the traditional op amp. This type of op amp requires local feedback (between its output and inputs) to make it stable. Under steady DC conditions, both the input terminals are virtually at the same voltage and this determines the output voltage setting. However, though both resistors of the voltage divider affect the DC level of the converter's output, from the AC point of view, only the upper resistor enters the picture. So the lower (R_b) is considered just a DC biasing resistor, and therefore we usually ignore it in control loop (AC) analysis.
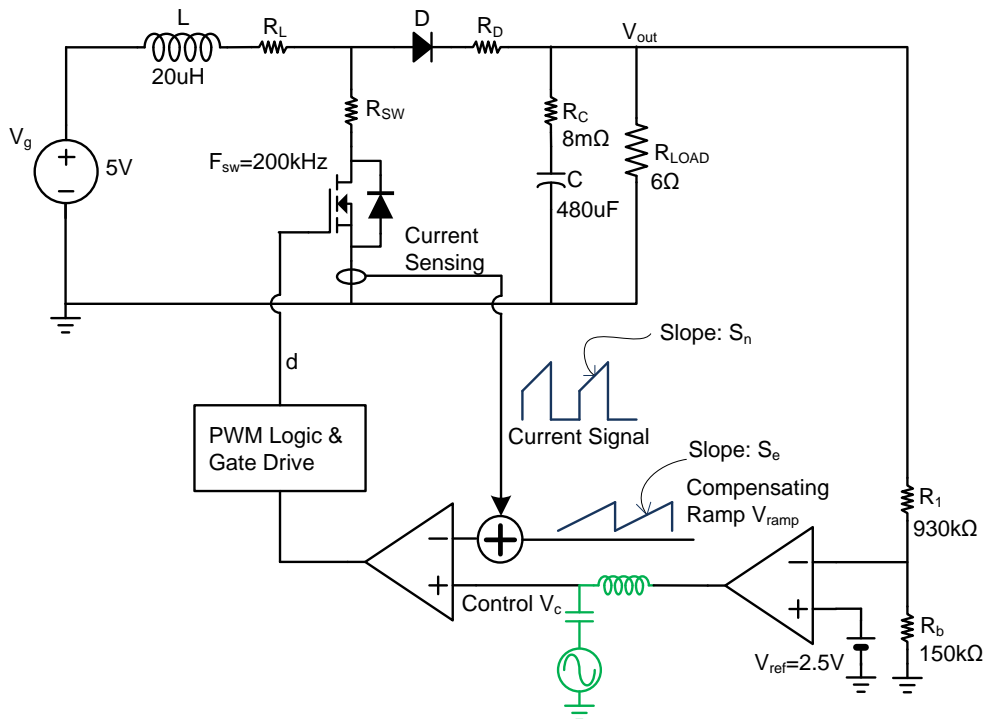
Figure 6.   Control-to-Output Transfer Function with Current-Mode Boost Converter

Figure 7 shows a comparison of the control-to-output for current-mode boost converter, and the control-to-output for voltage-mode boost converter. Note that the RHP zero is exactly the same as that for voltage-mode control. Using current-mode does not move this at all. The current-mode boost converter is easier to compensate, though, since we do not need to deal with the additional double pole response of the LC filter that is present with voltage-mode control.
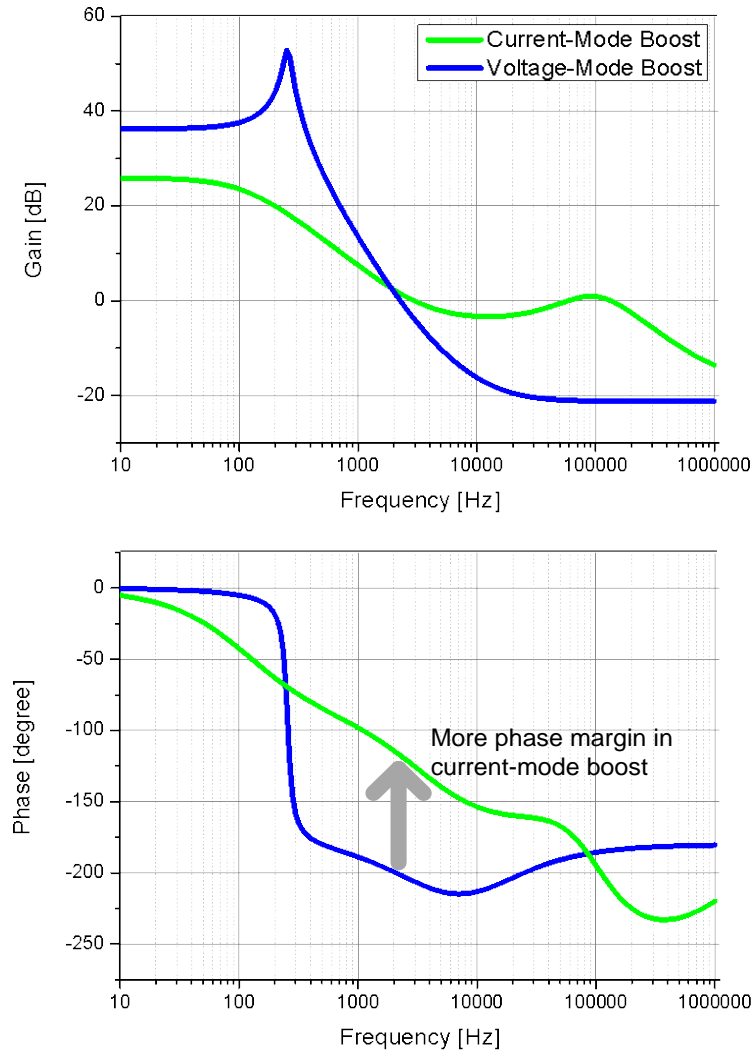

Figure 7.   Comparisons of Current-Mode and Voltage-Mode Control-to-Output Transfer Functions

4   Boost Converter (Current-Mode) Feedback Compensation

Now we are ready to design the feedback loop of current-mode boost converter understanding the control scheme. In order to control the boost converter, it is now necessary to design a feedback amplifier to compensate for the naturally-occurring characteristics of the power stage. Figure 8 shows a Type II compensation amplifier. This compensation scheme adds an RC branch to flatten the gain, and improve the phase response in the mid-frequency range. The increased phase is achieved by increasing the separation of the pole and zero of the compensation.
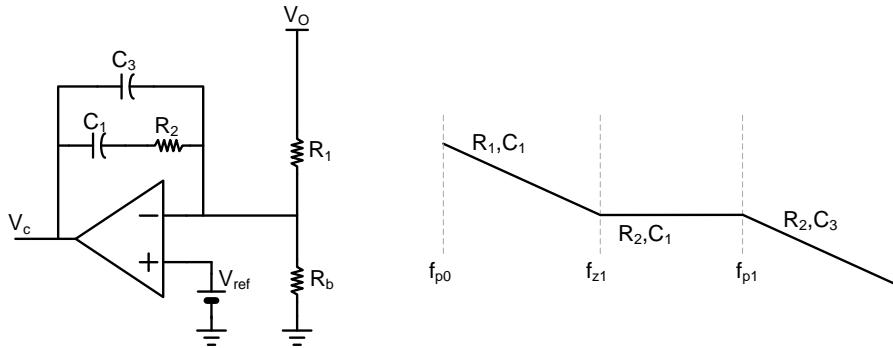
Figure 8.   Type II Compensator with Gain Curve

Note that this type of compensator still always has a net negative phase and it cannot be used to improve the phase of the power stage. For this reason, Type II compensators cannot be used for a voltage-mode boost converter where there is a large phase drop just after the resonant frequency, as shown in Figure 7. Type II compensators are usually reserved for current-mode control compensation, or for converters that always operate in the DCM region.

Type II (an origin pole, plus a pole/zero pair) gives us one pole-at-zero (fp0) and one pole (fp1) and one zero (fz1). We always need a pole-at-zero in the compensation for achieving high DC gain, good DC regulation, and low-frequency line injection. Note that four components (R_1, R_2, C_1, and C_3) are involved in determining the poles and zero, and the locations of the poles and zero are:

$$f_{p0} = \frac{1}{2\pi \times R_1 \times C_1} \tag{11}$$

$$f_{p1} = \frac{1}{2\pi \times R_2 \times C_3} \tag{12}$$

$$f_{z1} = \frac{1}{2\pi \times R_2 \times C_1} \tag{13}$$

and the transfer function (H(s)) for the feedback block with Type II is:

$$H(s) = \frac{(1+R_2 \times C_1 \times s)}{(R_1 \times C_1 \times s) \times (1+R_2 \times C_3 \times s)} \quad \text{if } C_1 >> C_3 \tag{14}$$

We can find the required C_1, R_2, and C_3 once we select R_1 with the desired f_{p0}, f_{p1} and f_{z1}.

$$C_1 = \frac{1}{2\pi \times R_1 \times f_{p0}} \tag{15}$$

$$R_2 = \frac{f_{p0} \times R_1}{f_{z1}} \tag{16}$$

$$C_3 = \frac{f_{z1}}{2\pi \times R_1 \times f_{p0} \times f_{p1}} \qquad (17)$$

The boost converter with current-mode control can operate successfully with just a Type II compensator and has four main characteristics. These are: a single pole at low frequency determined by the output capacitor & load resistor, an ESR zero and an RHP zero which moves with operating conditions. Also, there is a pair of double poles at half the switching frequency. Q is controlled with ramp addition.

In selecting values for the Type II compensator, these characteristics are taken into consideration in the placement of poles and the zero. The following list contains the design rules for the current-mode boost converter:

1. The first pole (f_{p0}) of the compensator is placed at the origin from an integrator.

2. The compensation zero (f_{z1}) is placed at one-fifth the selected crossover frequency.

3. The second pole (f_{p1}) of the compensator is placed coincident with the ESR zero or the RHP zero frequency, which is lower.

4. The crossover frequency should be less than about one-tenth the switching frequency.

5. The crossover frequency should be less than about one-fifth the RHP zero frequency.

Based on these rules, Figure 9 shows an example for the compensator which has an appropriate shape, and usually a good phase margin.
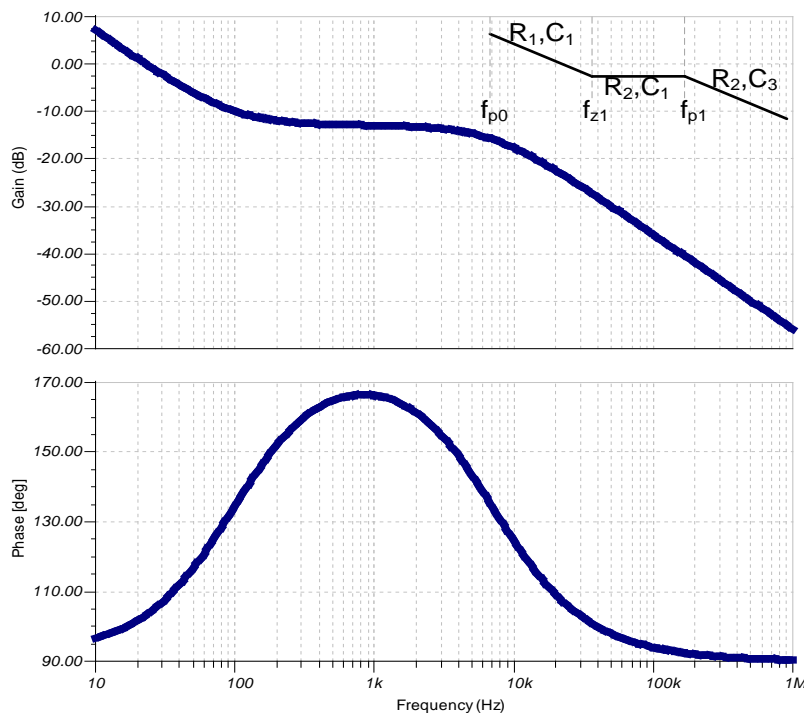
Figure 9.   Appropriate Compensator Design Example

As with the buck converter loop gain, it starts with a slope of –1, changes to –2 up to the compensation zero, then reverts to a –1 slope for the whole loop since this would compromise the low frequency gain.

5   Current-Mode Compensation Summary

Figure 10 and Figure 11 show the schematic and the loop gain of applying these rules to a boost converter example with the following design parameters (V_{in} = 5 V, V_{out} = 18 V, I_{out} = 3 A, L = 20 µH, F_{sw} = 200 kHz). The converter switches at 200 kHz. The crossover frequency is limited to about 0.6 kHz, passing it through with –1 gain slope, and the phase margin measured to be 75 degrees at this crossover frequency.
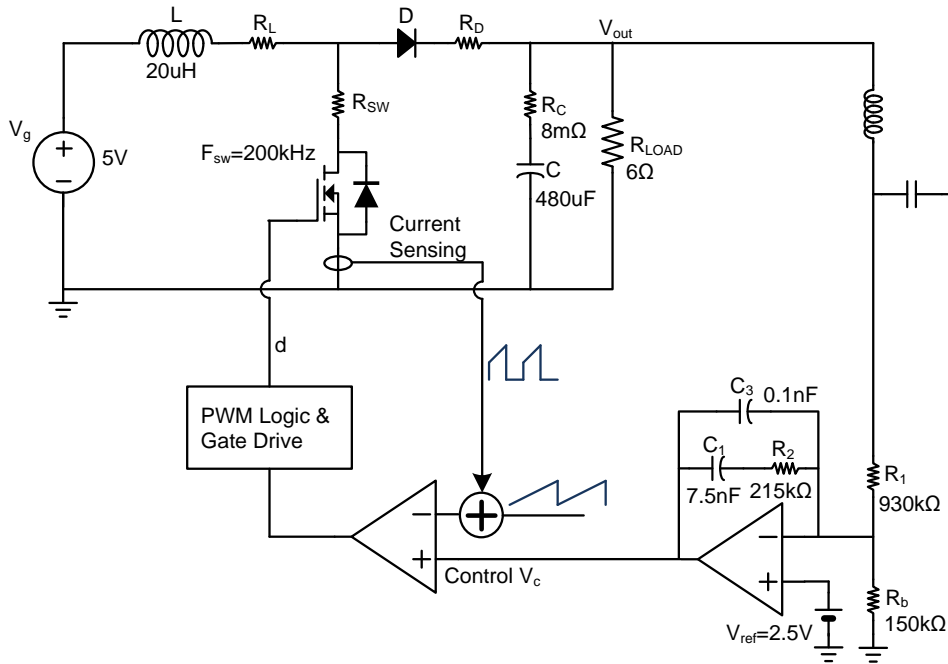
Figure 10.   Schematic with the Given Parameters

Figure 11 shows the resulting loop gain (in blue) with the compensation (in red) and the control-to-output (in green) waveforms when these rules are applied. The selected crossover frequency is 0.6 kHz, or one-fifth the RHP zero frequency (3.6 kHz).

Note that in the design rules, there was no requirement to cross the loop over in excess of the resonant filter frequency. This characteristic has already been eliminated by the current feedback loop. This is a very important observation, especially for current-mode boost converter which has a low-frequency RHP zero.

With voltage-mode control it is sometimes impossible to control such a converter with a loop crossover above the resonant frequency, and performance is very poor. Current-mode control solves this problem, and low RHP zero systems are controllable with good performance.
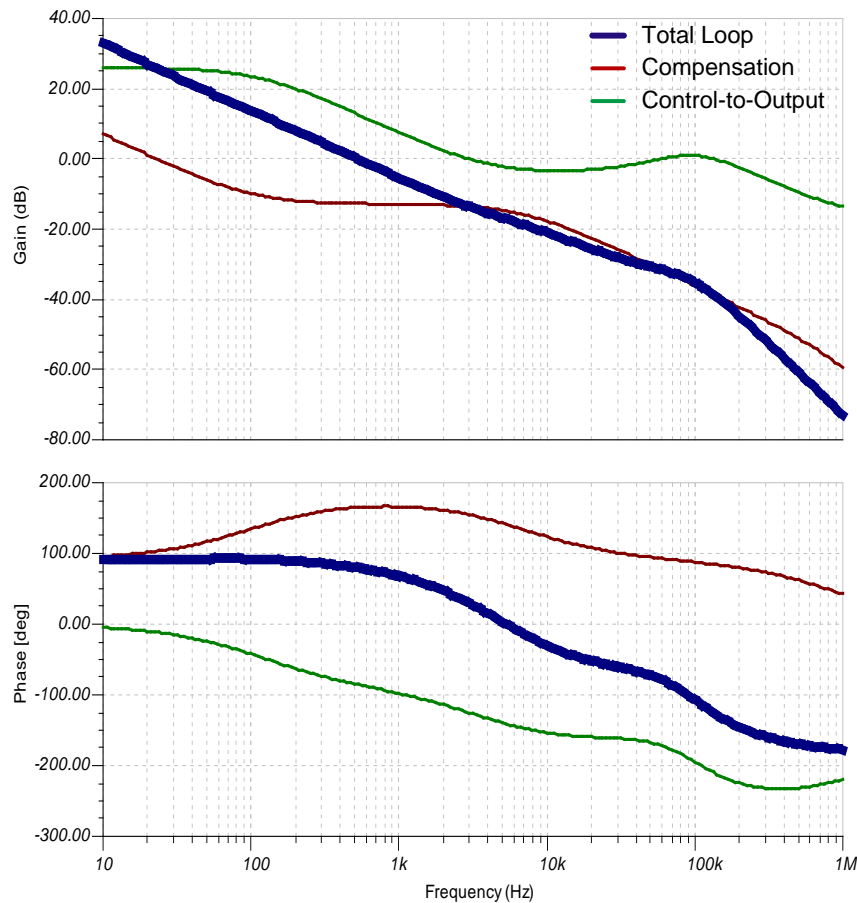
Figure 11.   Loop Gain and Phase Margin

6   Conclusion

Compensation for current-mode boost converter is much easier than voltage-mode boost converter, even if the RHP zero is at a low frequency. The Type II compensation has simple design rules, and good stability is usually achieved on the first attempt. There is no minimum requirement for the crossover frequency, so you can always make the system stable regardless of the RHP zero frequency. The current loop eliminates the ringing frequency of the filter, and good performance is achieved even with a relatively low crossover frequency on the voltage feedback loop. The proper ramp must be added to the current-mode boost converter to damp the subharmonic oscillations as shown in Figure 4.

7   Reference

1.   *Practical Feedback Loop Analysis for Voltage-Mode Boost Converter* (SLVA633), SW Lee, Texas Instruments, January 2014.

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